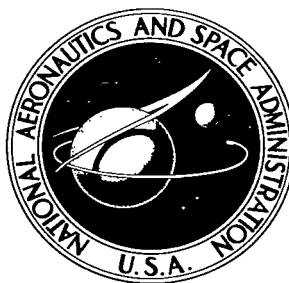


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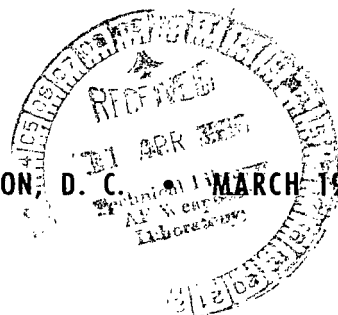
AN ULTRA-LOW-NOISE 1700-Mc PARAMETRIC AMPLIFIER SYSTEM

by P. H. Dalle Mura

*Goddard Space Flight Center
Greenbelt, Md.*

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

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ABSTRACT

As the use of solid-state varactor parametric amplifiers has expanded to include a variety of ultra-sensitivity receiving systems, technological improvement has succeeded in decreasing the noise temperatures of these devices to a point, even operating in an environment heated above room temperature, approaching the theoretically minimum uncooled noise temperature. The paper describes a completely field-operational ultra-low-noise parametric amplifier achieving, at room temperature and above, noise temperatures previously attainable only by lowering the varactor temperature significantly below room temperature.

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INTRODUCTION

During the past few years solid-state varactor parametric amplifiers have gained permanent status in a wide variety of ultra-sensitive receiving systems. Among these are ground and air based radars, tropospheric-scatter communication systems, and satellite tracking and data acquisition ground terminals. All of these systems demand the use of ultra-low-noise receivers coupled with low-noise antenna systems. The past few years have also seen the noise temperatures of varactor parametric amplifiers steadily decrease in step with the improvement in the technology of varactor fabrication. The dynamic quality factor of varactors has improved to the point where today the theoretically minimum uncooled paramp noise temperature is being approached in practice.

An ultra-low-noise parametric amplifier system employing very high quality varactor diodes was developed* for utilization at the Space Tracking and Data Acquisition Network terminals employing low-noise 85 foot antennas. The system contains five identical paramp channels having a common pump source. Each paramp is a single stage, one-port, negative-resistance type amplifier operating normally at a center frequency of 1705 Mc. The nominal gain is 20db and the 3db bandwidth is 15 Mc. For good stability, each amplifier employs a 4-port circulator, which is in essence an input isolator and a 3-port circulator combined into a single physical package, and a coaxial output isolator. An automatic power leveling circuit is used to maintain the pump power at a constant value over long periods of time.

Physically, the system consists of two enclosures; a microwave box containing the paramps and pump source, and an electronics box containing several power supplies. Each box provides electrical shielding and thermal insulation, the latter for the purpose of maintaining the interior components of each box at a constant temperature. Temperature control is necessary to insure high degrees of gain and phase stabilities.

Normally, the microwave box would be located at the antenna prime focal point, while the electronics box would be located at a remote point, possibly on the back side of the antenna

*The parametric amplifier system was developed under contract NAS5-3270 by Sperry Microwave Corp., Clearwater, Florida.

reflector. The microwave box occupies a volume of 2.2 cubic feet and weighs 100 pounds, while the electronics box has a volume of 1.6 cubic feet and weighs 70 pounds.

The major design objective of this program was the achievement of an operationally-packaged 5-channel parametric amplifier system displaying an input noise temperature of 75°K (1.0db noise figure) or less for each channel. It will be shown in this report that this objective has been achieved, and that the theoretically minimum noise temperature for an uncooled 1700 Mc amplifier pumped at 23.8 Gc is approached.

THEORY

One of the earliest definitions of varactor performance was set forth by Uhler, (Reference 1) wherein he proposed a diode figure of merit based on a "cutoff frequency," f_c , defined by

$$f_c = \frac{1}{2\pi R_s C_{min}} \quad , \quad (1)$$

where R_s = varactor series resistance, and C_{min} = varactor capacitance for reverse bias voltage just short of breakdown. This cutoff frequency is derived from the basic relationship for the Q of a varactor, namely

$$Q = \frac{1}{\omega R_s C} \quad , \quad (2)$$

where f_c is the point at which the maximum Q (occurring at C_{min}) becomes equal to unity.

This cutoff frequency is useful for comparing varactors of the same type or roughly comparing varactors of different types. However, since it contains no measure of the varactor performance under pumped conditions, it cannot be used as an absolute comparison of the quality of varactors employed in low noise parametric amplifiers.

A copious number of theories describing the operation of varactor parametric amplifiers have been developed which attempt to set forth the parameters determining the paramp noise performance. Among these are publications by Blackwell and Kotzebue (Reference 2), Penfield and Rafuse (Reference 3), and Mortenson (Reference 4) which define the paramp noise figure in terms of different varactor parameters. However, the results agree quite well, both theoretically and experimentally, even though the diode quality factor in each case is defined in a different manner to include the effect of pumping. The method of Penfield and Rafuse is straightforward and easily applied using known varactor parameters, therefore it will be employed in this analysis.

Penfield and Rafuse define the varactor cutoff frequency f_c by

$$f_c = \frac{S_{max} - S_{min}}{2\pi R_s} \quad , \quad (3)$$

where

R_s = varactor series resistance,

S_{max} = maximum varactor elastance = $\frac{1}{C_{min}}$,

and

S_{min} = minimum varactor elastance = $\frac{1}{C_{max}}$.

For most varactors, $C_{max} \gg C_{min}$, so that f_c reduces to $1/2\pi R_s C_{min}$ which is the same as Uhlir's definition given earlier in Equation 1. In order to account for the effects of pumping, a "pumped" figure of merit is defined as

$$Q_p = M_1 f_c, \quad (4)$$

where a modulation ratio M_1 whose value is determined by the variation of varactor elastance coefficients with pumping power,

$$M_1 = \frac{|S_1|}{S_{max} - S_{min}}. \quad (5)$$

The value of M_1 depends on both the form and the degree of pumping. With full, sinusoidal *voltage* pumping, the value of M_1 is 0.212 for an abrupt-junction varactor and 0.178 for a graded-junction varactor. With full, sinusoidal *current* pumping, the value of M_1 is 0.25 for an abrupt-junction varactor and 0.212 for a graded-junction varactor. As the degree of pumping decreases from full pumping, the value of M_1 decreases for both voltage and current pumping. Thus, to achieve the optimum in paramp performance (including lowest noise), the varactor must be pumped so that its elastance varies over the full range from forward conductance to reverse breakdown.

The basic noise temperature formula for a parametric amplifier as derived by Penfield and Rafuse is,

$$T = T_d \frac{f_s}{\beta f_i} \frac{(M_1 f_c)^2 + \beta^2 f_i^2}{(M_1 f_c)^2 - \beta f_s f_i} + T_i \frac{f_s}{f_i} \frac{R_i}{R_s + R_i} \frac{(M_1 f_c)^2}{(M_1 f_c)^2 - \beta f_s f_i}, \quad (6)$$

where

T_d = diode physical temperature,

T_i = idler termination physical temperature,

f_s = signal frequency,

f_i = idler frequency = $(f_{pump} - f_s)$,

$M_1 f_c$ = pumped quality factor,

R_s = diode series resistance,

R_i = idler termination resistance,

and

$$\beta = \frac{R_s + R_i}{R_s} .$$

The first term of Equation 6 is the noise contribution of the diode series resistance, R_s , while the second term is the contribution from the idler termination, R_i . In actual practice, the first term imposes a fundamental lower limit on the paramp noise since R_s can never be zero for a practical varactor, while the idler termination, R_i , can be made zero by terminating the varactor reactively at the idler frequency. Theoretically, the lower noise limit is achieved when $R_s = 0$, in which case Equation 6 reduces to

$$T = T_i \frac{f_s}{f_i} . \quad (7)$$

Equation 7 is often employed, especially in quick approximations. However, it can easily lead to serious errors, particularly in the design and analysis of very-low-noise parametric amplifiers.

For most well designed parametric amplifiers, including the subject of this report, it can be assumed that $R_i = 0$, so that Equation 6 reduces to

$$T = T_d \frac{f_s}{f_i} \frac{(M_1 f_c)^2 + f_i^2}{(M_1 f_c)^2 - f_s f_i} . \quad (8)$$

For a given f_s and varactor, there exists a finite pump frequency which minimizes the noise temperature. This optimum pump frequency is obtained by differentiating Equation 8 with respect to f_i and setting the result equal to zero. This produces the following expression for the optimum pump frequency

$$f_p (\text{optimum}) = \sqrt{(M_1 f_c)^2 + \omega_s^2} , \quad (9)$$

or since generally $M_1 f_c \gg \omega_s$,

$$f_p (\text{optimum}) \approx M_1 f_c . \quad (10)$$

Equation 8, the expression for noise temperature of the parametric amplifier with $R_i = 0$, is plotted in Figure 1, with noise temperature as a function of pump frequency for $f_s = 1.7$ Gc, $M_1 f_c = 60$ Gc, and T_d (diode temperature) = 290°K. It is clear that the curve displays a broad

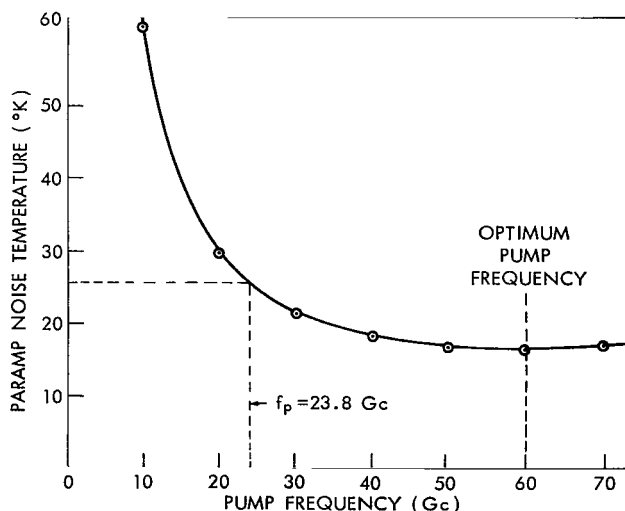


Figure 1—Noise temperature plotted as a function of pump frequency where $f_s = 1.7$ Gc, $T_d = 290^\circ\text{K}$, and $M_1 f_c = 60$ Gc.

frequency of 23.8 Gc was chosen as a good compromise between optimum noise performance and the availability of a stable, long-life klystron with sufficient power output to drive five paramps simultaneously.

The high quality varactor employed is a Sylvania D5047B gallium arsenide type having a typical cutoff frequency of 300 Gc at -6 volts bias. Assuming nearly full current pumping for this graded junction varactor, an $M_1 = 0.2$ is derived from curves given in Reference 3. Bearing in mind that the diode temperature is 330°K (135°F), due to the fact that the interior of the paramp enclosure is heated to maximum ambient, the theoretical noise contribution of the diode is computed from Equation 8 and turns out to be 30.6°K . To find the overall noise temperature of the amplifier, contributions from the input isolator-circulator and the input filter must be included. The input isolator-circulator has a loss of 0.3db from input port to paramp port, while the input filter has a loss of approximately 0.15db. Figure 2 depicts the overall noise temperature calculation which results in a theoretical parametric amplifier noise temperature equal to 69.4°K .

DESIGN

The parametric amplifier is designed to operate with the idler frequency much higher than the diode self-resonant frequency, and therefore an external idler cavity is required. Figure 3 is a sectional drawing showing the physical layout

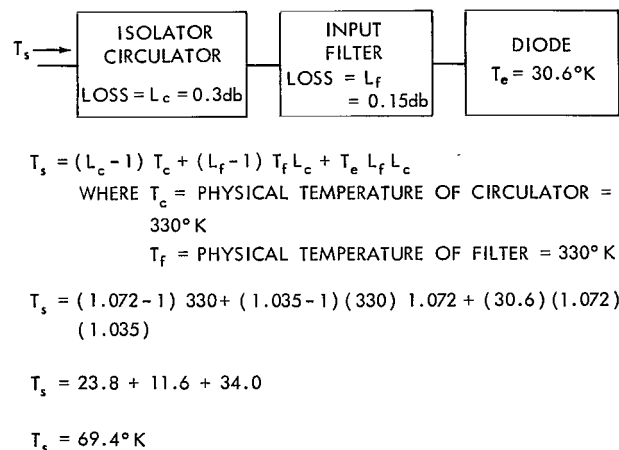


Figure 2—Overall noise temperature calculation.

minimum, therefore very little in noise temperature is sacrificed by pumping at frequencies as low as $0.33 M_1 f_c$. For our amplifier, a pump

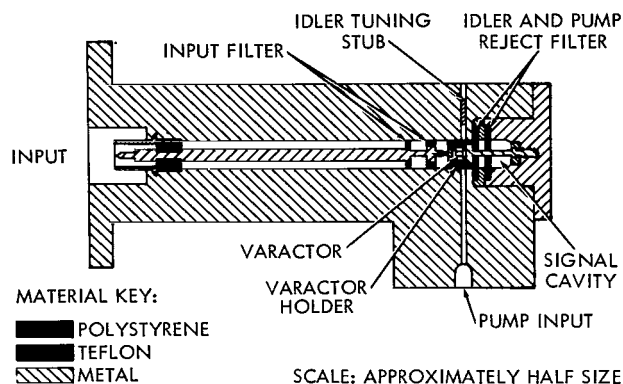


Figure 3—Paramp physical layout.

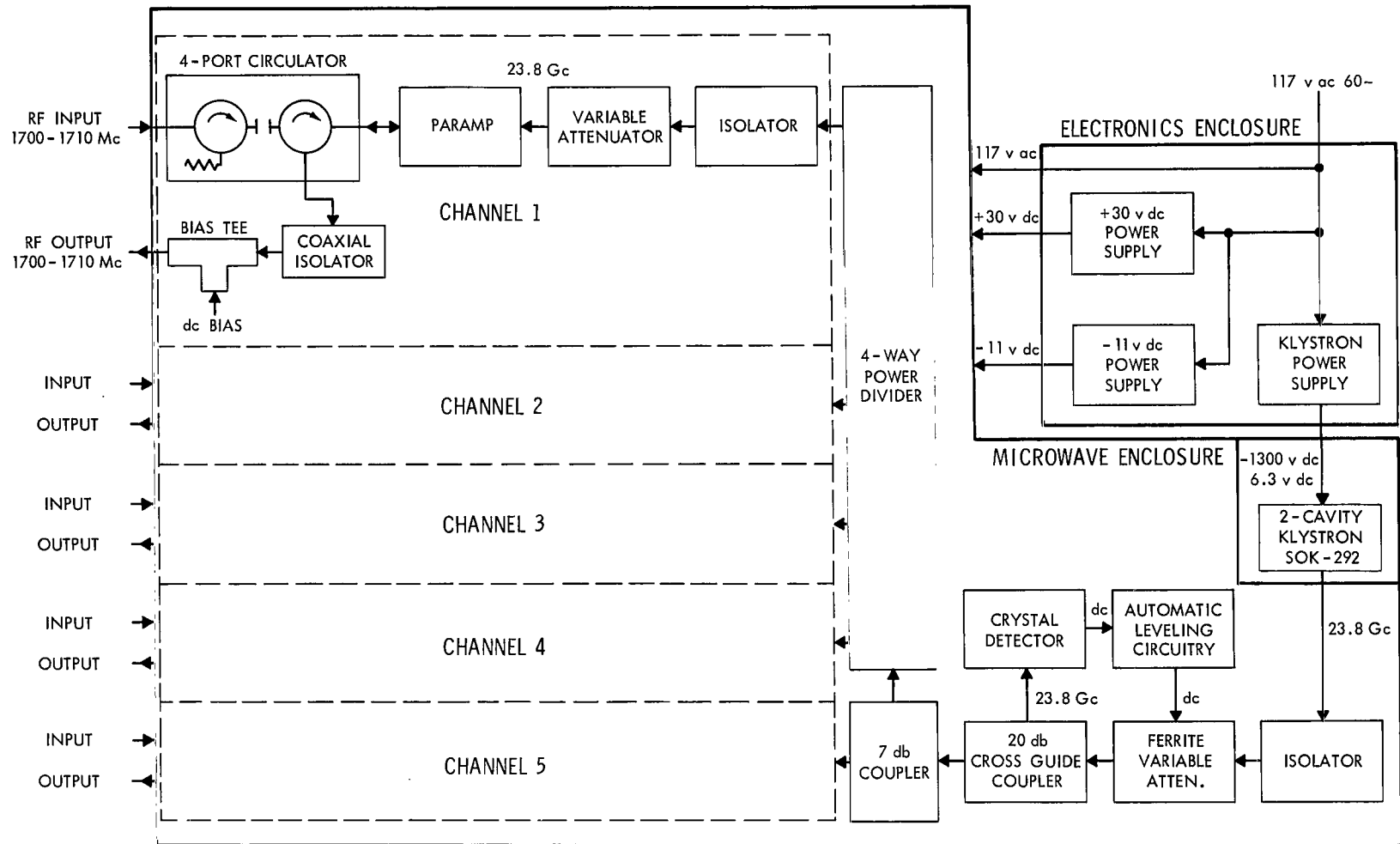


Figure 4—System block diagram.

of the main paramp components. The input line is coaxial, and the input connector is a modified type N connector which slips into a similar type on the circulator. The generator impedance required by the varactor for proper gain and minimum noise temperature is calculated to be approximately 50 ohms. Therefore the input transformer found in most parametric amplifiers is not required here. A coaxial disc-type low-pass filter is employed in the input line to prevent pump and idler signals from propagating out through the paramp input lines.

The diode is encapsulated in a cylindrical, polystyrene holder which supports the diode firmly in the paramp, as well as allowing its easy removal in case of failure. A part of the input filter and the entire signal tuning cavity are removed during the diode replacement operation. The diode is easily separated from these components, and the entire process can be accomplished in a matter of several minutes.

Pump and idler energy is blocked from the signal tuning cavity by a two section coaxial filter. Pump energy is fed into the diode by means of a reduced height and width waveguide which, acting as a waveguide below cutoff, blocks signal and idler energy from the pump line. The idler cavity was tuned experimentally for idler resonance in the breadboard model by means of an adjustable idler tuning stub (Figure 3); however it is fixed in the final amplifier.

The main amplifier block was accurately machined in halves and then precisely assembled by means of positioning pins and screws. All of the critical low loss microwave circuits are gold flashed to minimize surface deterioration.

Figure 4 is a block diagram of the entire parametric amplifier system. The pump source is a 2-cavity klystron which provides good stability and a power output of more than 1 watt at 23.8 Gc. The klystron is located within a separate temperature controlled enclosure which maintains the klystron output flange at approximately 100°C. Before the pump power is divided for distribution to each amplifier channel, it is passed through an automatic leveling system which maintains a constant pump power in spite of the gradual falloff of klystron power output due to aging. After the pump power is divided into five equal signals, each signal is then fed into a paramp through a ferrite isolator. By employing an isolator in each channel, the possibility of cross-coupling between channels through the pump circuitry is minimized. Each paramp channel contains a variable attenuator in the pump line and a variable dc varactor bias control to allow adjustment of the paramp tuning and gain.

The microwave enclosure is an insulated box whose outside dimensions are 18 inches by 16 inches by 13 inches. It occupies a volume of approximately 2.25 cubic feet. The walls are one inch thick polyurethane foam insulation in order to help maintain the interior at a constant temperature of +130°F. Figure 5 is a photograph of

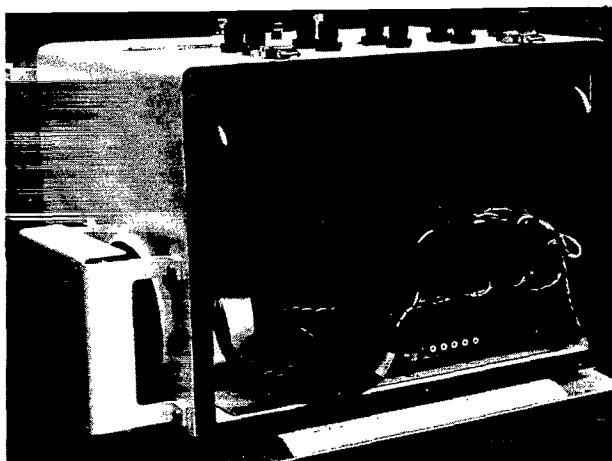


Figure 5—Paramp system microwave enclosure.

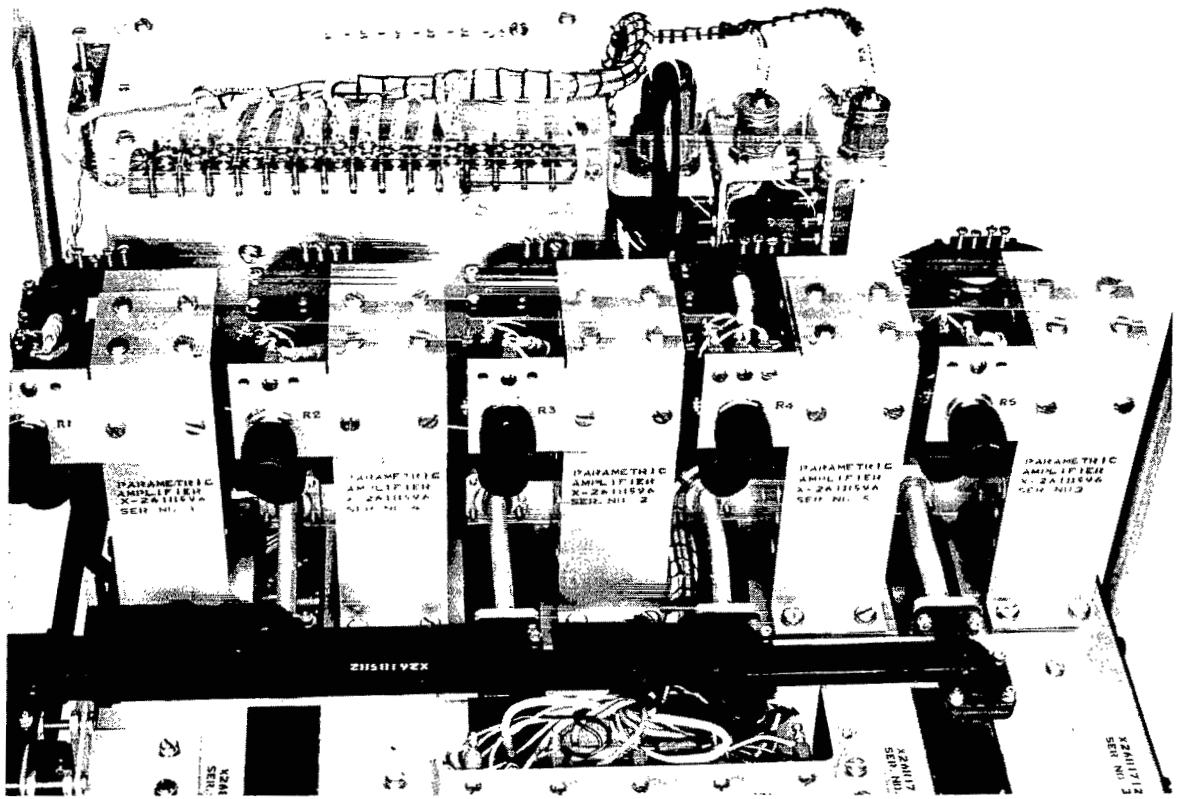


Figure 6—Paramp system microwave components.

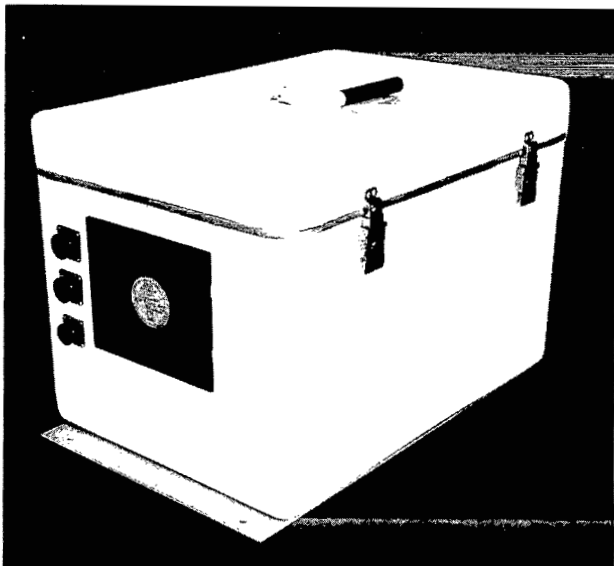


Figure 7—Paramp system electronics enclosure.

the microwave enclosure with the top removed, while Figure 6 shows the complete microwave system removed as a unit from the enclosure.

The electronics enclosure shown in Figure 7 is also fabricated of one inch thick polyurethane foam insulation and contains all of the system power supplies and the power distribution system. The +30 volts dc and -11 volts dc supplies are commercial transistorized supplies. The klystron power supply was designed by Sperry to deliver -1300 volts dc at 65 milliamperes for the klystron beam and +6.3 volts dc at 0.5 ampere for the klystron heater. The size of the electronics enclosure is 18 inches by 12 inches by 12 inches, and it occupies a volume of 1.5 cubic feet.

TEST PROGRAM

General Operating Characteristics

During the initial operation of the complete paramp system after receiving it from the contractor, it was discovered that the gain of each channel was affected drastically by intermittent interference generated within the system. It was ascertained that the sources of interference were two silicon controlled rectifiers (SCR's) located in the microwave enclosure, which are employed to control the klystron temperature and the enclosure temperature. These SCR's generated pulses when they triggered which traveled back through the ac power lines to the klystron high voltage supply. This supply also contains two back-to-back SCR's on the ac input side of the power transformers, their function being to provide regulation of the klystron high voltage output by controlling the ac input waveform. The intermittent triggers generated by the microwave enclosure SCR's were of sufficient amplitude to cause the klystron supply SCR's to also fire intermittently, the end result being a random variation of the klystron high voltage output. Since this high voltage supplies the beam voltage for the pump klystron, the pump power fed to each parametric amplifier displayed a drastic change in level in synchronization with the firing of the microwave enclosure SCR's. Thus the gains of all of the paramps are affected simultaneously.

It was decided that the best solution to the problem would be to redesign the klystron high voltage supply to immunize it against the offending SCR triggers, rather than try to eliminate the triggers themselves. It was reasoned that the former course would provide additional protection against external sources of interference which might enter the ac power line as well as against the internally generated triggers.

The entire high voltage power supply was redesigned at GSFC. The original supply employed two SCR's at the ac input to regulate against input voltage variations, and it used three cascaded transistors as a series regulator on the high voltage output. The redesigned supply eliminated both the SCR's and the transistor series regulators, and it substituted in their place a single 5881 tube employed as a series regulator driven by a four-transistor control circuit. The redesigned supply completely eliminated the susceptibility to any random triggering, and it now provides good regulation over a plus and minus ten percent change of the ac input voltage.

After correcting the power supply defect, it is observed that the paramps exhibit excellent gain stability, easy tuning by means of the bias and pump power controls, and a symmetrical gain response curve. It is possible to vary the gain continuously from 0 to 30db without inducing instability or self-oscillation. The system requires approximately two hours of warmup after cold turn-on to reach ultra-stable operation, however after the first thirty minutes the changes are rather small. Although the gain of each paramp is higher than normal at turn-on, none goes into oscillation.

Table 1 presents the results of performance characteristic measurements described in the remaining paragraphs of this section of the report.

Table 1

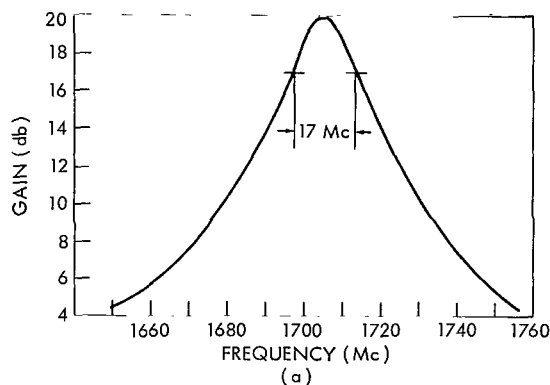
Performance Characteristics
(1700 Mc, 5-Channel Paramp).

Channel	Gain (db)	Frequency Center (Mc)	Bandwidth (Mc)	Noise Temperature (°K)	0.5db Gain Compression (dbm)	Input VSWR	Gain Change 5:1 Source VSWR (db)	Maximum Cross-Channel Coupling (db)	Gain Change 117V \pm 10% Input (db)
1	20	1706	15	70	-23	1.50	1.0	-55 to Ch. 3	0.2
	17	1704	21	66	-19	1.41	-		-
2	20	1705	16	81	-28	1.15	0.4	-77 to Ch. 1	0.2
	17	1704	22	70	-25	1.17	-		-
3	20	1705	16	77	-33	1.16	0.3	-80 to Ch. 1	0.0
	17	1703	22	78	-29	1.10	-		-
4	20	1706	16	74	-33	1.20	0.2	-57 to Ch. 5	0.1
	17	1704	22	89	-29	1.22	-		-
5	20	1705	17	68	-36	1.08	0.3	-76 to Ch. 4	0.1
	17	1704	24	72	-33	1.13	-		-

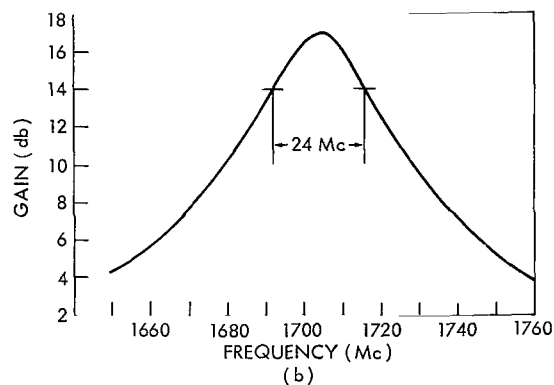
Gain Response

As mentioned previously, each paramp tunes easily and smoothly to any center frequency over the range from 1680 Mc to 1720 Mc by means of dc bias on the varactor and an attenuator in the pump power waveguide. The gain response was measured by means of an Alfred Sweep Generator, a crystal detector, and an oscilloscope. The gain was measured and set by the substitution method, in which the sweep generator output is first connected directly to the detector input to obtain a reference level, and then the paramp is inserted between the sweep generator and the detector. The amount the sweep generator output is reduced to again reach the reference level is the net gain of the paramp. The often-employed technique of turning off the pump power to set the reference level rather than removing the paramp from the circuit is incorrect, because the "cold" paramp losses are not taken into account. This leads to a net gain of several db less than the desired value.

Typical gain response curves (taken on amplifier number 5) are plotted in Figure 8. Each curve is an ordinary single-tuned response, because both the signal and idler cavities are single-tuned circuits. The center frequency gain is 20db and the 3db bandwidth is 17 Mc in Figure 8a. When the net gain is reduced to 17db, the bandwidth then increases to 24 Mc as shown in Figure 8b. Although the bandwidth is relatively narrow (generally characteristic of external idler cavity paramps), it adequately covers the 1700-1710 Mc satellite band.



(a) Gain = 20db



(b) Gain = 17db

Figure 8—Gain response curves.

Noise Temperature

The most important characteristic of any parametric amplifier is its low input noise temperature which allows it to amplify very weak signals without appreciably deteriorating the signal-to-noise ratio. The measurement of input noise temperature must be done carefully in order to avoid erroneous, and usually over-optimistic, results. The method employed here is the so-called "Y-factor technique" using a coaxial hot-cold load for the noise source.

The Y-factor is defined by the equation:

$$Y = \frac{T_h + T_e}{T_c + T_e} \quad (11)$$

where,

T_e = input noise temperature of device being tested,

T_h = temperature of hot load,

T_c = temperature of cold load.

Solving for T_e :

$$T_e = \frac{T_h - Y T_c}{Y - 1} \quad (12)$$

The Y-factor is determined experimentally by the use of a noise measurement setup as shown in Figure 9. A second paramp is used immediately following the paramp under test, in order to eliminate the effects of mixer noise contribution. The second paramp contributes less than one degree Kelvin of noise to the overall system noise, therefore the noise temperature measured is

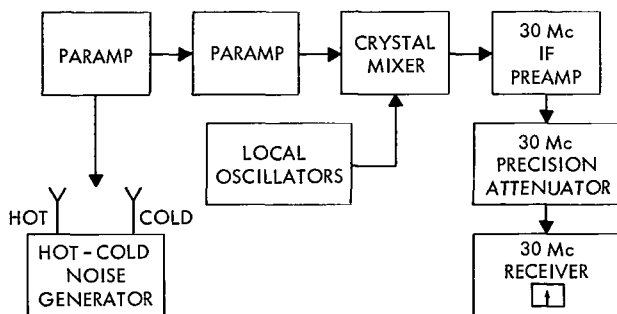


Figure 9—Noise measurement setup.

the true noise temperature of the first paramp stage alone. The second paramp also reduces the RF bandwidth and thereby makes the error due to image noise negligible.

The input of the paramp under test is first connected to the cold termination, and the receiver output meter is then set to a convenient reference point by means of either the preamp or receiver gain control. The reading of the precision attenuator is noted at this time. The input of the paramp is next connected to the hot

termination, and the attenuation of the precision attenuator is increased until the same reference reading is obtained on the receiver output meter. The Y -factor is the difference between the two readings of the precision attenuator; or more basically it is the difference between the noise output with the hot-load noise source and the noise output with the cold-load noise source. For the noise generator employed (AIL Type 70, Hot-Cold Body, Standard Noise Generator), the hot load is maintained at 373°K , and the cold load is maintained at 77.3°K (liquid nitrogen). Thus, knowing T_h , T_c and Y , the input noise temperature, T_e , of the paramp can be computed using Equation 12.

There are two corrections which must be applied to the noise measurement outlined above. First, the cold-load temperature, T_c , for the AIL Noise Generator is generally accepted as being several degrees higher than the specified value. According to measurements made at Bell Telephone Laboratories and at Microwave Physics Corp., a reasonable estimate of the cold-load temperature at its output jack is 83°K . Therefore, in computing T_e , 83°K is used for T_c instead of 77.3°K . The second correction is a 0.05db loss incurred in a type N-to-Type GR adapter required to connect the noise generator to the paramp input terminal. This loss introduces a noise contribution of 3.5°K which must be subtracted from the measured noise temperature of the paramp.

The results of the noise temperature measurements are shown in Table 1. The errors described above have been taken into account, and therefore the noise temperatures presented are corrected values. For 20db gain, the value of T_e ranges from 68°K up to 81°K . The accuracy of the measurement techniques is only $\pm 0.1\text{db}$ (approximately $\pm 7^{\circ}\text{K}$), therefore it is evident that within this error of measurement, each amplifier meets the noise temperature requirement of 75°K or less.

Earlier in this report, it was calculated that the theoretical noise temperature should be 69.4°K , including circulator and input filter losses. It is apparent from the measured data that this theoretical figure has been approached, and actually equalled for paramps number 1 and number 5. The slightly higher noise temperatures of amplifiers 2, 3 and 4 are probably the result of slightly higher losses in the circulator and input filter than those assumed.

Input VSWR

Paramps which are employed in monopulse tracking receivers are required to have input VSWR's of approximately 1.5 to 1 or lower. Since the paramps designed here are single-port,

reflection-type devices, a high input VSWR can result unless adequate isolation is designed into the input circuit. The most reliable method of insuring low input VSWR is to provide a low-loss ferrite isolator in the input circuit immediately preceding the input circulator. This method is not susceptible to critical tuning or matching requirements, and in addition it provides additional gain and phase stabilities for the amplifier. Its only serious drawback is an increase in input noise temperature of approximately 11°K due to its 0.15db insertion loss.

The input VSWR of each amplifier was measured by the conventional slotted line technique with the amplifier operating at both 17db and 20db of gain. The VSWR was 1.25 to 1 or lower for all channels except channel number one. The VSWR for channel one was less than 1.5 to 1. The results are displayed in Table 1.

Dynamic Range

The dynamic range of a low-noise amplifier is a measure of its ability to provide linear amplification for input signals varying over a wide range of amplitudes. Amplification is linear from noise threshold up to signal levels strong enough to cause overloading of the amplifier input circuit. The paramps were checked for dynamic range by observing the output signal as the input signal was increased from a low level up to the point where compression becomes evident. When the relationship between output signal level as a function of input signal level deviated from a straight line by 0.5db, the upper limit of dynamic range was reached.

Table 1 shows the results of the dynamic range, or gain compression, measurement. As can be seen from the data, 0.5db compression occurs for 20db amplifier gain at roughly 3db lower than for 17db amplifier gain, as would be expected for an amplifier of this type. For the worst case (Channel 5), the specified compression occurred at -36dbm which is more than adequate for the signal levels to be expected in operational use.

Gain Stability

Each paramp channel was tested for gain stability when the input VSWR changed from a matched condition up to 5 to 1. The test setup used to make this test is depicted in Figure 10. By varying the variable attenuator, the input termination presented to the paramp can be varied from a VSWR of 1 (30db of attenuation) up to a VSWR of infinity (0db of attenuation). A precise VSWR can be obtained by use of the easily derived relationship,

$$L = \left(\frac{P - 1}{P + 1} \right)^{1/2} \quad (13)$$

where P = voltage standing wave ratio, and L = voltage attenuation ratio. Thus, for VSWR = 5, L = 1.8db (ratio 0.817). Since the directional coupler had a directivity of 30db, the attenuation changes in front of the coaxial short had a negligible effect on the amplitude of the signal from the HP 8614 signal generator.

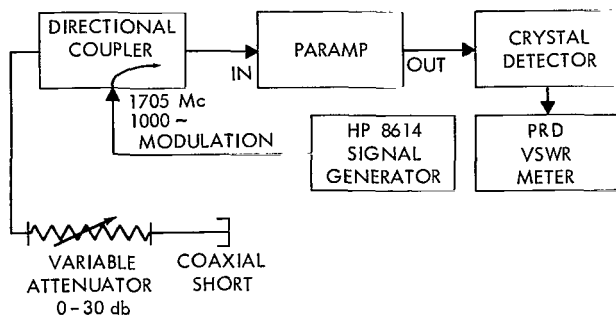


Figure 10—Test setup for gain stability versus input VSWR change.

times each day. Typical laboratory environmental conditions prevailed during the course of this test. The data are plotted in graphical form in Figure 11. It can be seen that the absolute gains of all five paramp channels remained within a $\pm 1.0\text{db}$ range for the entire period. This excellent gain stability can be attributed in great part to the stable pump power output brought about by the use of a closed-loop, automatic power leveling circuit in the pump output line. The critical nature of the effect of pump power variations on paramp gain stability is seen by examining the following empirical equation which relates these two quantities:

$$\Delta G = (\sqrt{G} - 2) \Delta P \quad (14)$$

where

G = paramp gain in power ratio,

ΔP = pump power change in db,

ΔG = gain change in db.

For $G = 20\text{db}$, $\Delta G = 8\Delta P$; therefore a change in pump power of 0.1db results in a gain change of 0.8db . An interesting point is that by reducing the paramp gain to 10db , $\Delta G = 1.16\Delta P$. Therefore two 10db gain amplifiers in cascade, having a total gain of 20db , would have a $\Delta G_{\text{total}} = 2.32\Delta P$, or a gain stability of almost four times greater than a 20db gain, single-stage paramp. This could be a practical approach for systems requiring very high degrees of gain stability.

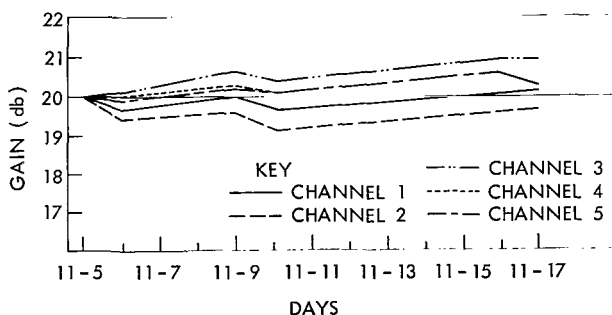


Figure 11—Long term gain stability (measured at 1200 hours).

The results of the test are shown in Table 1. The worst channel is number 1, where a change of 1.0db occurred. The other four channels displayed a gain change of less than 0.4db . Since Channel 1 also has the highest input VSWR (1.5 to 1), it can be assumed that the input isolator in Channel 1 is not providing as much isolation as the other four input isolators.

A measure of the long term, absolute gain stability was determined by operating the paramp system continuously for twelve days while checking the gain of each paramp channel several

The system was checked for gain stability as a function of ambient temperature by operating it inside of a temperature test chamber for three days. The temperature was cycled several times between -40°F and $+120^{\circ}\text{F}$, while the

absolute gain was monitored. At the start, each amplifier was set for 20db net gain. During the course of the temperature variations, the gain of each amplifier stayed within a range of +2db to -1db of the initial gain.

The gain variation with ambient temperature is more than adequate for STADAN monopulse tracking applications because of the following two factors: (1) the -40°F to +120°F temperature excursion is much more severe than would ever be encountered under actual operating conditions, and (2) post-comparator amplitude variations produce only a second order effect in amplitude monopulse tracking systems.

Gain stability for input power variation was measured by varying the 117 volts ac input power over a $\pm 10\%$ range and noting the change in gain of each channel. The results, shown in Table 1, indicate that the maximum change in gain for any channel was 0.2db.

Phase Stability

Amplitude monopulse tracking systems require that the relative phase between preamplifier output signals remain fairly constant to prevent the degradation of tracking accuracy. Field operational paramp systems have the requirement that the maximum phase shift between the two worst channel pairs over a 12 hour period be 10 electrical degrees.

Relative phase stability measurements were performed using a Wiltron Model 302 Phase Comparison System, which consists of a slotted line, with dual probe crystal detectors, and an electronic phase indicator unit. The two signals to be compared are fed into opposite ends of the slotted line and the probe carriage is adjusted for a null (zero center meter reading). Any phase shift between the two signals is read directly in degrees on the meter, with full scale deflection variable in steps from 0.6° up to 90°.

The results of the measurement of relative phase stability between channels is depicted in Figure 12. Channel 5 was used as the reference with all other channels compared to it. Over a 24 hour period, the worst relative phase shift amounted to four degrees between channels 2 and 5. Except for channel 2, all channels remained within one and one-half degrees of each other. This degree of phase stability is more than adequate for STADAN monopulse tracking applications.

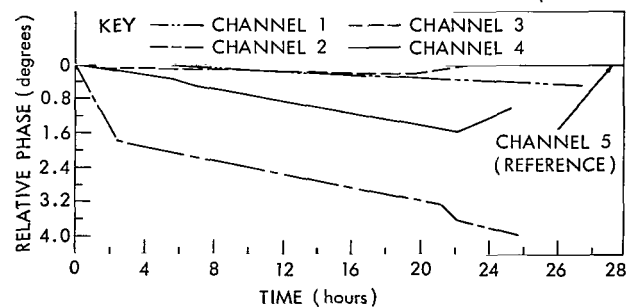


Figure 12—Relative phase stability.

Cross-Channel Coupling

Cross-channel coupling in amplitude monopulse preamplifiers can cause serious tracking errors to occur in the system. The isolation between channels must be at least 50db to reduce this

effect to a negligible quantity. The cross-channel isolation was measured by feeding a signal into the input of one channel, and using the output of that same channel as a reference, measuring the signal level at the output of the other four channels and comparing them with the reference. All unused input and output terminals were terminated with 50 ohm loads during these measurements. This measurement was repeated in turn for input signals into each of the other four channels with the results shown in Table 1. It is seen that the worst cross-coupling occurs from channel 1 to channel 3, and it is -55db down, well within the -50db specification.

CONCLUSIONS

An ultra-low-noise parametric amplifier system which advances the state of the art has been described. Completely field-operational paramps were developed which give noise temperatures achieved previously only by means of lowering the varactor temperature significantly below room ambient temperature. Conversely, these paramps operate in a temperature environment which is heated *above* room temperature to insure good stability regardless of the operating temperature environment. This adds approximately 9°K to the noise temperature which would be achieved if the system were operating at room temperature.

It has been shown that the noise temperature achieved (75°K) approaches closely the theoretical noise temperature possible for a paramp operating at a signal frequency of 1.7 Gc, a pump frequency of 23.8 Gc, and employing a varactor having a pumped figure-of-merit of 60 Gc.

In addition to exhibiting outstanding noise performance, the paramp system provides multi-channel operation in a completely operational package, which lends itself to prime focus mounting in any low-noise, large reflector, tracking and data antenna system. The gain and phase stabilities realized are well within the basic requirements for amplitude monopulse tracking systems. The paramp system is easily tuned, readily serviceable, including varactor diode replacement in a matter of minutes, and it employs a long life, stable, and rugged two-cavity klystron pump source.

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